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Eleventh day of February 2002

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Patent Specification

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Inventor: Neville Thiele

Title: Improved crossover networks and method

Applicant Techstream Pty Ltd Australia, ACN 062 162 458

Assignee Techstream Pty Ltd

References

Patent	Inventor	Title
US5568560	Combest	Audio crossover circuit
US5377274	Meyer et al	Correction circuit and method for improving the transient behavior of a two way loudspeaker system
US5373653	Kukurudza	Self damping speaker matching device
US5359664	Steuben	Loudspeaker system
US5327505	Kim	Multiple output transformers network for sound reproducing system
US5185801	Meyer et al	Correction circuit and method for improving the transient behavior of a two way loudspeaker system
US5153915	Farella	Speaker filtering circuit and support therefore
US5109423	Jacobson et al	Audio system with amplifier and signal device
US4897879	Geluk	Multi-way loudspeaker system
US4882760	Yee	Sound reproduction system
US4771466	Modafferi	Multi-driver loudspeaker apparatus with improved crossover filter circuits
US4769848	Eberbach	Electroacoustic network
US4723289	Schreiber et al	Stereo electroacoustic transducing
US4691362	Eberbach	Dihedral loudspeakers with variable dispersion circuits
US4653103	Mori et al	Loudspeaker structure and system
US4638505	Polk et al	Optimised low frequency response of loudspeaker systems having main and sub-speakers

US4606071	White Jr	Loudspeaker system utilising an equaliser circuit
US4597100	Grodinsky	Ultra high resolution loudspeaker system
US4593405	Frye et al	Loudspeaker system with combination crossover and equaliser
US4589135	Baker	Zero phase shift filtering
US4583245	Gelow et al	Speaker system protection circuit
US4578809	Eberbach	Dihedral loudspeakers with variable dispersion circuits
US4483015	Strohbeen	Compensation network for loudspeakers
US4475233	Watkins	Resistively damped loudspeaker system
US4430527	Eberbach	Loudspeaker crossover delay equalisation
US4429181	Freadman	Audio system
US4426552	Cowans et al	Speaker distortion compensator
US4421949	Eberbach	Electroacoustic network
US4410063	Yasue et al	Loudspeaker system
US4383134	VonRecklinghausen	Loudspeaker systems
US4349697	Skabia	Sound reproduction system
US4348552	Siccone	Direct/reflecting speaker system and triangular shaped enclosure
US4348549	Berlant	Loudspeaker system
US4340778	Cowans et al	Speaker distortion compensator
US5315102	Eberbach	Speaker crossover networks
US4295006	Tanaka et al	Speaker system
US4287389	Gamble	High fidelity speaker system
US4282402	Liontonia	Design of crossover network for high fidelity speaker system
US4249042	Orban	Multi-band cross-coupled compressor with overshoot protection circuit
US4243840	Kates	Loudspeaker system
US4238774	Iwahara	Frequency band dividing filter using delay line filter
US4237340	Klipsch	Crossover network for optimising efficiency and improving response of loudspeaker system
US4229619	Takahashi et al	Merthod and apparatus for driving a multi-way speaker system
US4229618	Gamble	High fidelity speaker with negative feedback
US4198540	Cizek	Compensated crossover network
US4179669	Dobson et al	Amplifying and equalising
US4154979	Barker	Woofer efficiency
US4133975	Barker III	Loudspeaker system with broad image source with directionality control for the tweeter
US4100371	Bayliff	Loudspeaker system with phase difference compensation

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US3838215	Haynes Jr	Speakers and crossover circuits
US3657480	Cheng et al	Muti-channel audio system with crossover network feeding separate amplifiers for each channel with direct coupling to low frequency loudspeaker
US3457370	Boner	
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Improved crossover networks and method

Field of invention

The present invention relates to the effective splitting of a source signal into separate frequency bands for downstream handling by equipment dedicated to the respective frequency bands with the intent that the recombination of the signals in any domain has the desired amplitude and phase integrity.

Crossovers are most commonly used in loudspeaker systems so that the electroacoustic transducers may be dedicated to appropriate frequency bands. In such applications the crossover design may be active or passive or a combination thereof, and may be placed before, within or after the amplifiers feeding the electroacoustic transducers.

The invention overcomes many of the performance limitations in prior art crossovers and in particular the phase and amplitude integrity and the rate of cut off across the crossover transition region.

Background art

It is usual to provide more than one electroacoustic transducer in a loudspeaker system in order to provide for the best compromise between performance and cost. As each Electroacoustic transducer is designed to provide optimum performance over a range of frequencies in particular, it is necessary to provide some filtering of the applied signals covering the bands of more than one electroacoustic transducer in order to split the signals into the desired bands. Many designs of such filters have been described in the art and references have been listed elsewhere in this document.

What is desired for the crossover is that the resulting filter or crossover provides signals to the two or more drivers which, when recombined through the electroacoustic transducers provide an accurate recreation of the original signal, taking into account the amplitude response of the drivers and the nature of the radiation pattern of the whole loudspeaker into the listening environment.

Common shortcomings in the prior art have been inability to achieve a flat

amplitude response across the one or more crossover frequencies, inability to control the phase of the signals across the crossover region, inability to roll off the response of each electroacoustic transducer quickly enough particularly at the low frequency side of the crossover frequency in order to avoid out of band signals introducing distortion, or causing damage to the electroacoustic transducers in question.

Summary of the invention

This invention comprises a new class of crossover characterised by the inclusion of null responses at frequencies close to the crossover region. For each crossover transition the high pass and low pass outputs have amplitude responses that go through nulls at frequencies either side of the transition region but which add together to produce a combined output that is flat across the whole spectrum.

Benefits of such a configuration are improved amplitude response, improved phase response and improved rate of signal attenuation close to the corner frequency for each band.

The invention is characterised mathematically in the ideal case by identical denominators of paired low pass and high pass functions whilst the numerators comprise terms that in each individual filtering function are all of even or all of odd order, thereby maintaining constant phase difference between low and high pass outputs. The sum of the numerator pairs has the same squared modulus as their common denominator thereby maintaining the sum of the combined outputs as a constant giving a flat amplitude response across the whole spectrum.

The invention may be further characterised by a notch associated with a band stop filter adjacent to the or each corner frequency.

The invention may be further characterised by a rate of roll-off of response between the corner frequency and the notch or band stop frequency for each transition region that is controlable by the location of the or each notch or band stop frequency relative to the corner frequency.

Figure 1 shows the generalised form of the frequency response for the

invention where F_{NL} is the lower null centre frequency for the low pass filter, F_{NH} is the upper null centre frequency for the high pass filter, F_{PEAKL} is the lower peak frequency for the low pass filter, , F_{PEAKH} is the upper peak frequency for the high pass filter, F_{INNERL} is the highest frequency at which the output of the low pass filter equals the peak value below the null for the low pass filter, F_{INNERH} is the lowest frequency at which the output of the high pass filter equals the peak value above the null for the high pass filter and F_X is the crossover frequency.

Embodiments

The invention is realised as pairs of low pass and high pass filters. The invention can be realised in circuits of different orders depending on the desired rate of rolloff of the resultant crossover. The invention can be realised using passive, active or digital circuitry for the filter realisation as is known in the art, or combinations thereof. Combinations include but are not limited to an active low pass and passive high pass filter pair of any desired order, digital filter low pass and active high pass of any desired order, passive low pass and passive high pass of any desired order, Digital low pass and digital high pass of any desired order and active low pass and digital high pass filter realisations.

The invention can be further realised wherein the filter response is produced with a combination of electrical and mechano-acoustic filtering as would be the case where the loudspeaker driver and/or enclosure design realised part of the filter response according to the invention described above. In such an embodiment, the order of filter and realisation in passive, active or digital circuitry would equally apply.

Figure 2 shows the schematic circuit diagram for a sixth order active circuit embodiment of the invention. In this figure the characteristic elements of the invention are shown for the low pass filter utilising IC2, IC3 and IC4. An inverter, IC1 is provided between the low and high pass sections to provide the correct phase for the signals. IC4 realises the notch in the low pass filter utilising the Sallen Key topology known in the art. IC7 realises the notch in the high pass filter

also utilising the Sallen & Key topology known in the art. By this means the appropriate nulls in response are generated. IC3 generates the required second order filter transfer function for the low pass filter and IC2 generates the two single order cascaded section response as required. IC6 generates the required second order filter transfer function for the high pass filter and IC5 generates the two single order cascaded section response as required. These filter sections use the Sallen & Key topology known in the art. The outputs of IC4 and IC7 provide the required signals to the low and high frequency loudspeakers. Inspection of the signals in this network will reveal the response curves shown in figures 3 and 4.

Figure 3 Shows the resultant low pass amplitude and phase response for the circuit in figure 2.

Figure 4 Shows the resultant high pass amplitude and phase response for the circuit in figure 2.

Figure 4 shows the resultant summed output phase and amplitude response for the circuit in figure 2.

The invention is further described in the attached paper "Loudspeaker crossovers with notched responses" by the inventor. This paper is to be presented at the Audio Engineering Society (AES) conference January 2000, Paris France.

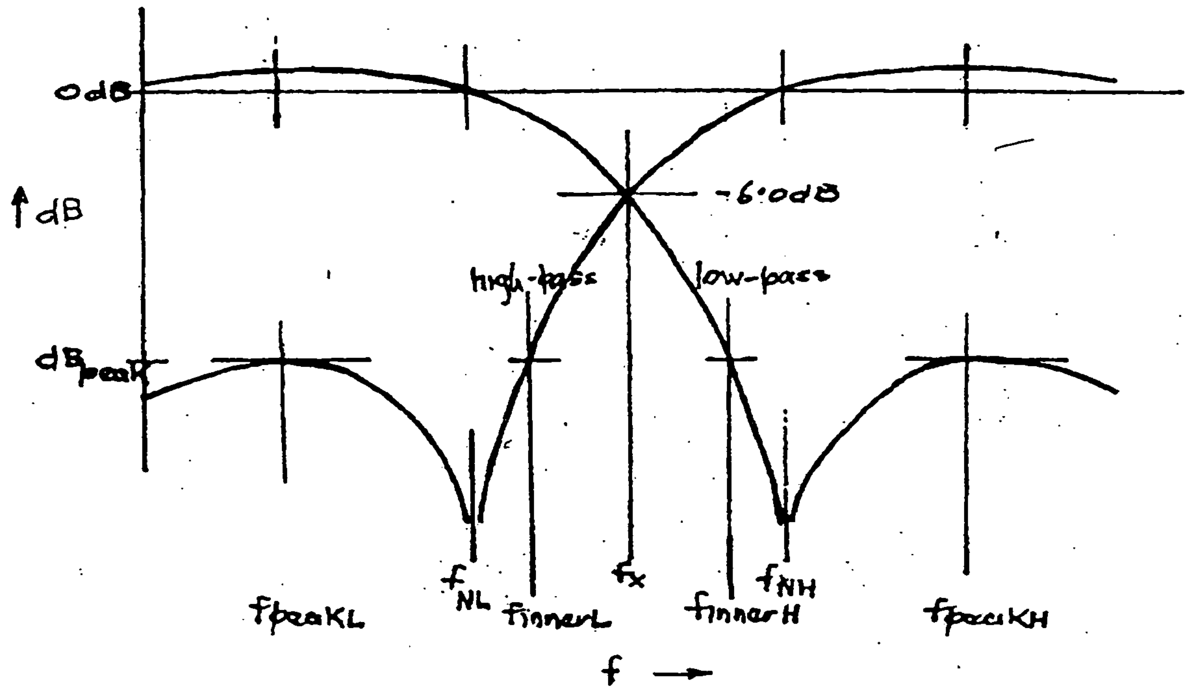

Figure 1: Generalised amplitude response of the invention

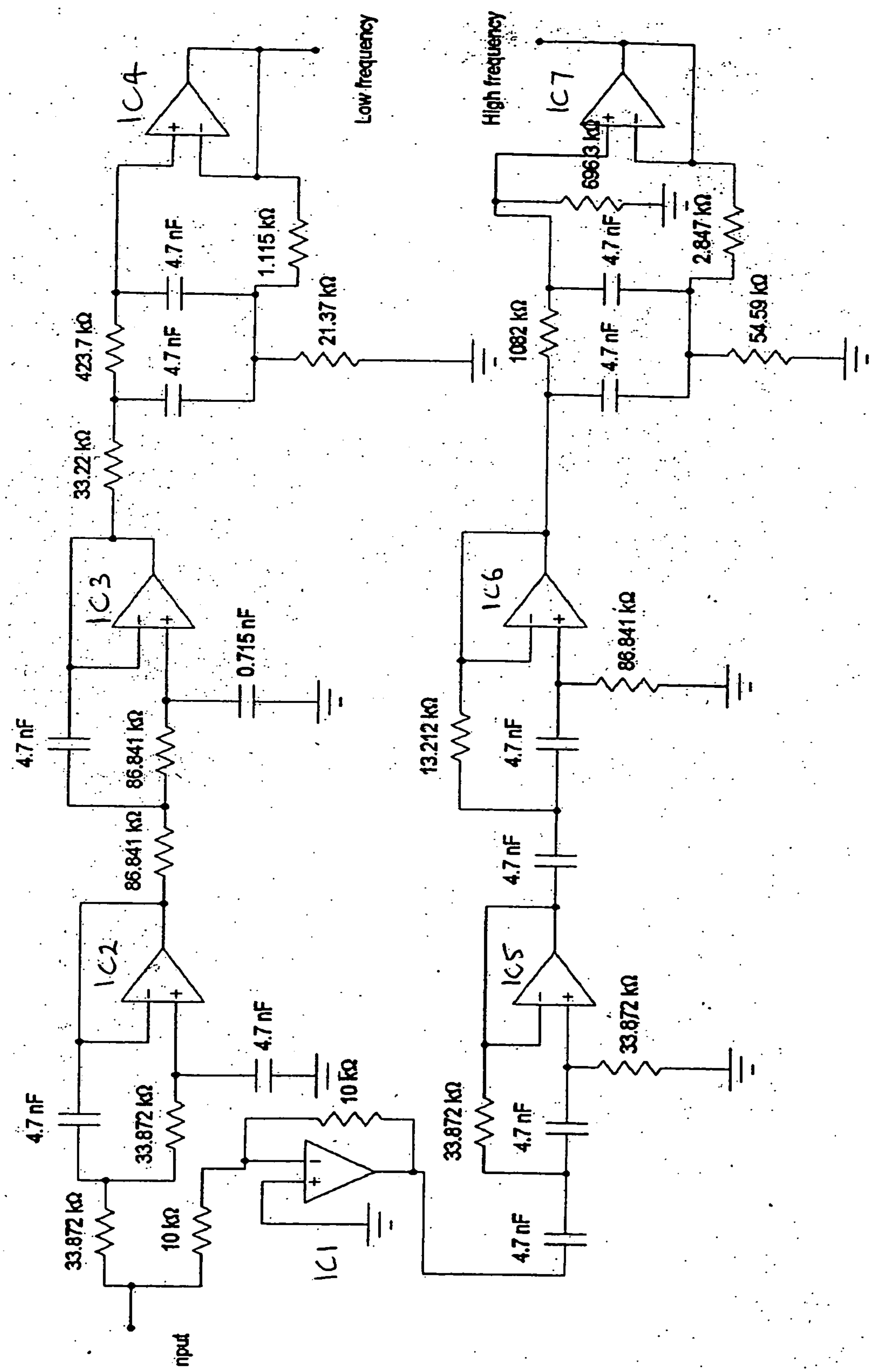

Figure 2 Circuit schematic for a sixth order active embodiment of the invention

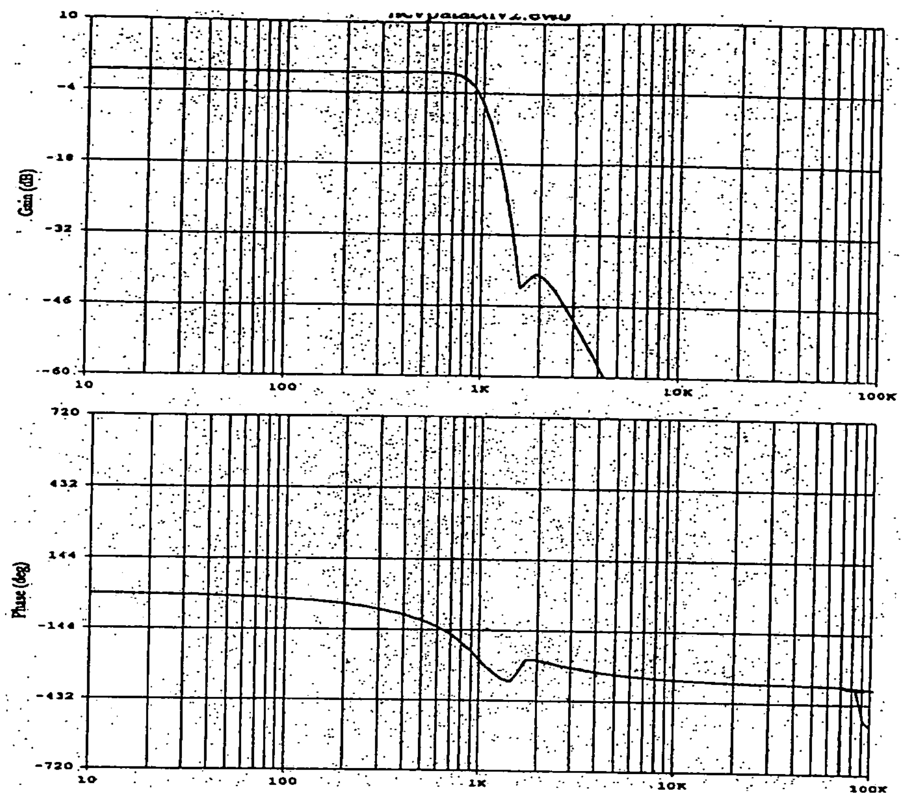

Figure 3 Lowpass phase and amplitude response for the circuit of figure 2

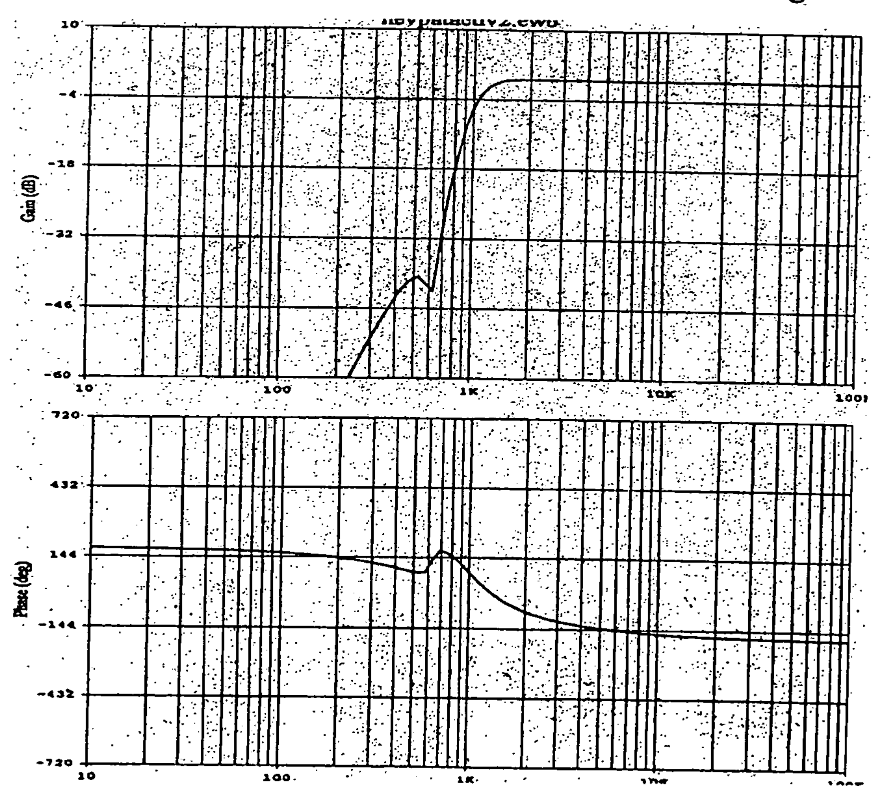

Figure 4 High pass phase and amplitude response for the circuit of figure 2

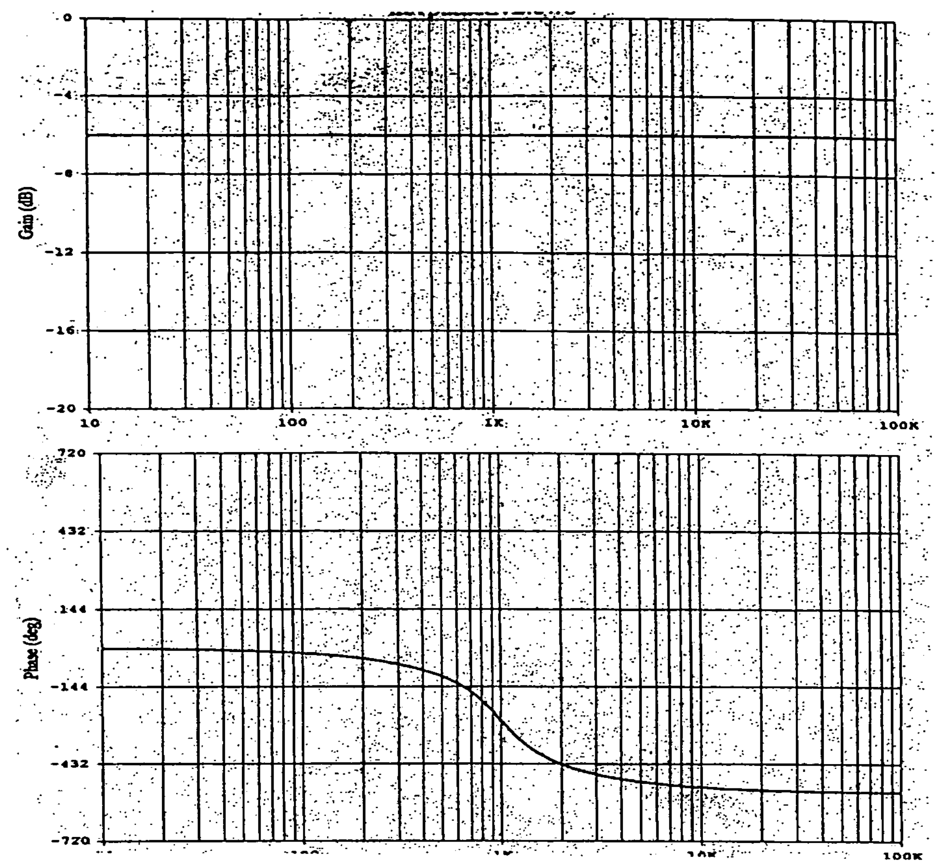

Figure 5 Resultant output phase and amplitude response for the circuit of figure 2

A paper "Loudspeaker crossovers with notched responses" by the inventor and further describing the invention is attached. This paper is to be presented at the AES conference 2000, Paris France.

LOUDSPEAKER CROSSOVERS WITH NOTCHED RESPONSES

Summary. A class of crossover systems is described, which produce null responses at frequencies close to the transition, crossover, frequency, while the sum of their outputs, high pass plus low-pass, has an ideally flat response. When the nulls are moved to extreme frequencies, the notched functions degenerate into Butterworths, for odd order, and Linkwitz-Rileys, for even order. Active and passive realisations are presented.

This material is the subject of patent applications.

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Introduction The paper describes crossover systems for loudspeakers, in which filters separate electrical signals covering the whole audio spectrum into two or more ranges for feeding separate drivers. One highly important property of such systems is that the acoustical outputs from the drivers add together to produce a combined output with an, ideally, "flat" response of amplitude vs. frequency.

The paper describes what is believed to be a novel class of crossover transfer functions. It defines pairs of functions, high-pass and low-pass, whose individual outputs have amplitude responses that go through nulls at frequencies symmetrically either side of the transition, crossover, frequency f_X but add together to produce a combined output that is flat across the whole spectrum.

The denominators of the high-pass and low-pass functions are identical while the numerators, comprising terms that in each individual filtering function are either all of even order or all of odd order, ensure that a constant phase difference is maintained between the high-pass and low-pass outputs. The sum of the two numerators, high-pass plus low-pass, has the same squared modulus as their common denominator, so that the response of the combined outputs is flat across the whole spectrum. It is in fact an all-pass function.

This kind of response was suggested by a paper by Hardman[1], in which the summed response rippled within 1dB. That crossover however turns out, however, to be an approximation to a general class of functions, of any order, whose summed response is genuinely flat across the spectrum. The notch provides a greatly increased initial slope for a filter of a given order. Beyond the notch, the stop-band response of each filter rises again to a maximum amplitude that can be selected by the designer. Thus a balance can be struck between the initial attenuation slope and a height of response at the following maximum that contributes negligibly to the summed output.

General Form of the Crossover Function The transfer function of the summed output of the proposed crossover of nth order is

$$F(sT)_{\Sigma n} = \frac{\overset{\text{LOW-PASS}}{(1 + k^2s^2T_X^2)} \pm \overset{\text{HIGH-PASS}}{s^{n-2}T_X^{n-2}(k^2 + s^2T_X^2)}}{F_{DEN}(sT_X)} \qquad - (1)$$

where k is the ratio of lower notch frequency f_{NL} in the high-pass response to the crossover, transition, frequency f_X

$$k = f_{NL}/f_X = f_X/f_{NH} \qquad - (2)$$

where f_{NH} is the higher notch frequency in the low-pass response, and T_X is the associated time constant of the crossover frequency ($T_X = 1/2\Pi f_X$).

The common denominator is derived from the numerator of the summed response by factorising it into first and second order terms, changing the signs of any negative first order terms to positive and then re-multiplying all the factors together. The result is thus an all-pass response whose numerator is the product of all the factors with negative first order terms.

The generalised notched responses of the even-order functions are shown in Fig 1. Within the pass band of each filter, the in-band response rises at first to a small peak at the frequency of the out-of-band peak of the other filter. It then falls back to reference 0dB level at the other filter's notch frequency, and onwards to -6.0dB at the transition frequency f_X.

The response falls to a null at its f_N, then rises to dB_{PEAK} at f_{PEAK} before falling away again at extreme frequencies at a rate, for an nth order filter, of 6(n-2)dB per octave. The effective limit of its response is at f_{INNER} where it has fallen first to dB_{PEAK}.

The outputs of the complementary, high- and low-pass, filters are completely in phase between the frequencies of the two notches. At frequencies beyond, on either side, their outputs are completely out-of-phase. Thus each in-band response must rise a little, never more than 0.27dB for out-of-band peaks lower than -30dB, to compensate for the subtraction of the out-of-band output of the other filter.

The responses of the odd-order functions are similar to those of even order, except that, because the individual high- and low-pass outputs combine in quadrature, each is now down to -3.0dB, instead of -6.0dB, at the crossover frequency f_X. The individual outputs now have a constant phase difference of 90° at frequencies between the two notches. At frequencies beyond, the inversion of polarity leaves the two outputs to still add in quadrature. Thus the in-band responses now *fall* initially, by less than 0.01dB, before rising to reference level and then falling again to the stop band, in the manner of odd order elliptic function filters.

It turns out, not surprisingly, that when k is zero, so that the notch frequencies move outwards to zero and infinite frequencies, the transfer functions degenerate into Butterworths for odd order functions and double Butterworths (Linkwitz-Rileys) for the even order functions.

The group delay responses are similar to the "parent" response of the same order, with a somewhat lower insertion delay at low frequencies and a somewhat higher peak delay at a frequency below the transition f_X, as can be seen in Tables 1, 2 and 3, before diminishing towards zero at very high frequencies. This will become clearer from examining specific examples.

Even-Order Functions. We deal first with the even order responses which like their "parent" Linkwitz-Riley responses, are more forgiving than the odd-order, Butterworth, responses of frequency and phase response errors in the drivers, and have better directional "lobing" properties.

Second Order Response. There are no useful second order responses.

Fourth Order Response. The high-pass and low-pass outputs are combined by addition.

$$F(sT)_{\Sigma 4} = \frac{\overbrace{(1 + k^2s^2T_X^2)}^{\text{LOW-PASS}} + \overbrace{s^2T_X^2(k^2 + s^2T_X^2)}^{\text{HIGH-PASS}}}{F(sT)_{DEN4}} \qquad - (3)$$

$F(sT)_{DEN}$ is derived by factorising the numerator

$$F(sT)_{NUM4} = 1 + 2k^2s^2T_X^2 + s^4T^4$$

$$= [1 + sT_X\sqrt{\{2(1 - k^2)\}} + s^2T_X^2][1 - sT_X\sqrt{\{2(1 - k^2)\}} + s^2T_X^2] \qquad - (4)$$

The equivalent minimum-phase function for $F(sT)_{DEN4}$ is thus

$$F(sT)_{DEN4} = [\,1 + x_4 sT_X + s^2T_X^2\,]^2 \qquad - (5)$$

where

$$x_4 = \sqrt{\{2(1 - k^2)\}} \qquad - (6)$$

from which the individual low-pass and high-pass functions are

$$F(sT)_{LP4} = \frac{1 + k^2s^2T_X^2}{[\,1 + x_4 sT_X + s^2T_X^2\,]^2} \qquad - (7)$$

and

$$F(sT)_{HP4} = \frac{s^2T_X^2\,(k^2 + s^2T_X^2)}{[\,1 + x_4 sT_X + s^2T_X^2\,]^2} \qquad - (8)$$

and the summed response is the all-pass function

$$F(sT)_{\Sigma 4} = \frac{1 - x_4 sT_X + s^2T_X^2}{1 + x_4 sT_X + s^2T_X^2} \qquad - (9)$$

When k becomes zero, the responses degenerate to 4th order Linkwitz-Riley functions. The generalised notched responses are plotted in Fig. 1, and the values for the fourth order responses are shown in Table 1 in terms of a crossover frequency f_X of 1000 Hz. The height of the peak amplitude following the notch is dB_{peak}. In the bottom row of Table 1, figures for group delay response of the Linkwitz-Riley response $k = 0$ are shown for comparison. Also the frequencies dB_{40}, dB_{35} and dB_{30}, where the Linkwitz-Riley response is down 40dB, 35dB and 30dB respectively, replace f_{peakL}, f_{NL} etc.

It will be seen how steepness of the initial attenuation slope can be traded for magnitude of the following peak.

Table 1. Fourth Order Responses. Peak dB, Out-of-Band Frequencies(Hz) & Group Delays(μs) for various values of k

k^2	dB_{peak}	f_{peakL}	f_{NL}	f_{innerL}	f_X	f_{innerH}	f_{NH}	f_{peakH}	*Insertion Delay*	*Peak Delay*	*Gp at Hz*
1/3	30.4	414	577	633	1000	1580	1732	2415	368	613	796
1/4	35.7	355	500	550	1000	1820	2000	2818	390	589	759
1/5	39.7	316	447	491	1000	2037	2236	3162	403	577	741
0		317	367	425	1000	2352	2726	3154	450	543	644
		dB_{40L}	dB_{35L}	dB_{30L}		dB_{30H}	dB_{35H}	dB_{40H}			

for HP — in μs

The responses at f_X are -6.02dB for all values of k. The group delay figures for other frequencies of f_X can be scaled inversely with frequency from those quoted above.

Sixth Order Responses. The sixth order functions are derived in a manner similar to the fourth order response. As in sixth order Linkwitz-Riley response, the high-pass and low-pass outputs are combined by subtraction.

$$F(sT)_{\Sigma 6} = \frac{\overbrace{(1 + k^2 s^2 T_X^2)}^{\text{LOW-PASS}} - \overbrace{s^4 T_X^4 (k^2 + s^2 T_X^2)}^{\text{HIGH-PASS}}}{[(1 + sT_X)(1 + x_6 sT_X + s^2 T_X^2)]^2} \quad \text{- (10)}$$

where $x_6 = \sqrt{(1 - k^2)}$ - (11)

and the summed response is the all-pass function

$$F(sT)_{\Sigma 6} = \frac{(1 - sT_X)(1 - x_6 sT_X + s^2 T_X^2)}{(1 + sT_X)(1 + x_6 sT_X + s^2 T_X^2)} \quad \text{- (12)}$$

Table 2. Sixth Order Responses. Peak dB, Out-of-Band Frequencies(Hz) & Group Delays(μs) for various values of k

k^2	dB_{peak}	f_{peakL}	f_{NL}	f_{innerL}	f_X	f_{innerH}	f_{NH}	f_{peakH}	*Insertion Delay*	*Peak Gp Delay*	*at Hz*
0.5480	30.0	617	740	779	1000	1283	1351	1622	532	1146	930
0.4653	35.0	565	682	719	1000	1391	1466	1771	555	1075	915
0.3915	40.0	515	626	660	1000	1515	1598	1940	567	1025	901
0		465	512	565	1000	1769	1951	2151	637	873	818
		dB_{40L}	dB_{35L}	dB_{30L}		dB_{30H}	dB_{35H}	dB_{40H}			

Eighth Order Responses Again the eighth order functions are derived in a manner similar to that for the earlier functions. The low-pass and high-pass outputs are summed by addition.

$$F(sT)_{\Sigma 8} = \frac{\overbrace{(1 + k^2 s^2 T_X^2)}^{\text{LOW-PASS}} + \overbrace{s^6 T_X^6 (k^2 + s^2 T_X^2)}^{\text{HIGH-PASS}}}{[(1 + x_{81} sT_X + s^2 T_X^2)(1 + x_{82} sT_X + s^2 T_X^2)]^2} \quad \text{- (13)}$$

where $x_{81} = [\{(4 - k^2) + \sqrt{(8 + k^4)}\} / 2]^{1/2}$ - (14)

and $x_{82} = [\{(4 - k^2) - \sqrt{(8 + k^4)}\} / 2]^{1/2}$ - (15)

and the summed response is the all-pass function

$$F(sT)_{\Sigma 8} = \frac{(1 - x_{81} sT_X + s^2 T_X^2)(1 - x_{82} sT_X + s^2 T_X^2)}{(1 + x_{81} sT_X + s^2 T_X^2)(1 + x_{82} sT_X + s^2 T_X^2)} \quad \text{- (16)}$$

Table 3. Eighth Order Responses. Peak dB, Out-of-Band Frequencies(Hz) & Group Delays(μs) for various values of k

k^2	dB_{peak}	f_{peakL}	f_{NL}	f_{innerL}	f_X	f_{innerH}	f_{NH}	f_{peakH}	*Insertion Delay*	*Peak Gp Delay*	*at Hz*
0.6628	30.0	719	814	843	1000	1186	1228	1392	710	1761	965
0.5906	35.0	675	769	797	1000	1255	1301	1483	727	1643	956
0.5224	40.0	632	723	750	1000	1333	1384	1581	742	1558	949
0		652	606	563	1000	1534	1651	1776	832	1244	888
		dB_{40L}	dB_{35L}	dB_{30L}		dB_{30H}	dB_{35H}	dB_{40H}			

Odd Order Responses. In the same way as the "parent" Butterworth responses, the high-pass and low-pass outputs, which add in quadrature, can be summed either by addition or subtraction for a flat overall response. However, the group delay error is lower when the 3rd and 7th order outputs are subtracted and when the 5th (and 9th) order outputs are added.

Third Order

$$F(sT)_{\Sigma 3} = \frac{\overbrace{(1 + k^2 s^2 T_X^2)}^{\text{LOW-PASS}} - \overbrace{sT_X (k^2 + s^2 T_X^2)}^{\text{HIGH-PASS}}}{[(1 + sT_X)(1 + x_3 sT_X + s^2 T_X^2)]} \qquad - (17)$$

$$= \frac{1 - sT_X}{1 + sT_X} \qquad - (18)$$

where $$x_3 = 1 - k^2 \qquad - (19)$$

Fifth Order Responses.

$$F(sT)_{\Sigma 5} = \frac{\overbrace{(1 + k^2 s^2 T_X^2)}^{\text{LOW-PASS}} + \overbrace{s^3 T_X^3 (k^2 + s^2 T_X^2)}^{\text{HIGH-PASS}}}{(1 + sT_X)(1 + x_3 sT_X + s^2 T_X^2)} \qquad - (20)$$

$$= \frac{(1 - x_{52} sT_X + s^2 T_X^2)}{(1 + x_{52} sT_X + s^2 T_X^2)} \qquad - (21)$$

where $$x_{51} = [-1 + \sqrt{(5 - 4k^2)}]/2 \qquad - (22)$$

and $$x_{52} = [+1 + \sqrt{(5 - 4k^2)}]/2 \qquad - (23)$$

Seventh Order

$$F(sT)_{\Sigma 7} = \frac{\overbrace{(1 + k^2 s^2 T_X^2)}^{\text{LOW-PASS}} - \overbrace{s^5 T_X^5 (k^2 + s^2 T_X^2)}^{\text{HIGH-PASS}}}{(1 + sT_X)(1 + x_{71} sT_X + s^2 T_X^2)(1 + x_{72} sT_X + s^2 T_X^2)(1 + x_{73} sT_X + s^2 T_X^2)} \qquad - (24)$$

$$= \frac{(1 - sT_X)(1 - x_{72} sT_X + s^2 T_X^2)}{(1 + sT_X)(1 + x_{72} sT_X + s^2 T_X^2)} \qquad - (25)$$

The x coefficients of the factors of the seventh order numerator are found from the roots of the equation

$$x_7^3 - x_7^2 - (2 - k^2)x_7 + (1 - k^2) = 0 \qquad - (26)$$

Of the three roots the largest and the smallest magnitudes x_{71} and x_{73} are positive. The middle magnitude root is negative, and its sign is changed to positive to produce x_{72}. Thus for example, when $k^2 = 0.5$, the roots of the equation are +1.7071, -1.0000 and +0.2929, so the coefficients x_{71}, x_{72} and x_{73} are 1.7071, 1.000 and 0.2929 respectively.

Typical results for the odd order responses are not tabulated because they are believed to be of less interest than the even order responses.

Special Uses of the Notched Crossovers In the notched crossovers, the initial slope of attenuation is greatly increased over that of an un-notched filter of the same order, and the minimum out-of-band attenuation can be chosen by the designer, 30dB, 35dB, 40dB or whatever. However the attenuation slope is eventually reduced by 12dB per octave at extreme frequencies. The group delay error is also increased somewhat through never as much as that for the un-notched filter two orders greater.

These functions should be specially useful when crossovers must be made at frequencies where one or other driver, assumed to be ideal in theory, has an amplitude and phase response that deteriorates rapidly out-of-band, a horn for example near its cut off frequency. Another application is in crossing over to a stereo pair from a single sub-woofer, whose output must be maintained to as high a frequency as possible so as to minimise the size of the higher frequency units, yet not contribute significantly at 250Hz and above where it could muddy localisation.

Realising the Filters From the designer's point of view, the crossovers are most easily realised as active filters, with each second order factor of the transfer functions realised in the well-known Sallen and Key configuration. The exception of course is the one factor which provides the notch, with a transfer function of the form, for the low-pass filter

$$F(sT) = \frac{1 + qs(kT_X) + s^2(kT_X)^2}{1 + xsT_X + s^2T_X^2} \qquad - (27)$$

and for the high-pass filter

$$F(sT) = \frac{1 + qs(T_X/k) + s^2(kT_X/k)^2}{1 + xsT_X + s^2T_X^2} \qquad - (28)$$

where q is ideally zero and x is the coefficient appropriate to one factor of the desired denominator, e.g. $x_4 = \sqrt{2}(1 - k^2)$ for the factors of the fourth order crossover.

While q may be made zero in active filters using cancellation techniques, which depend on the balance between component values, quite small values of q can be realised in a Sallen and Key filter that incorporates a bridged T network[2]. Unless as a deep notch is really necessary, it will often be sufficient to let the notch "fill up" with a finite value of q.

In the sixth order notched crossover, for example, when the height of out-of band peaks are 30dB, 35dB and 40dB, then figures for q , which would be equivalent to 1/Q in passive components, of 0.16, 0.14 and 0.10 respectively ensure that the attenuation at the erstwhile notch frequency is no less than at the erstwhile peak and that there is no significant change in response at neighbouring frequencies.

Component values are tabulated in Table 4 for the network of Fig 2 to realise the function

$$F(sT) = \frac{1 + x_N sT_N + s^2T_N^2}{1 + x_D sT_D + s^2T_D^2} \qquad - (29)$$

Table 4 Component Values for Sallen & Key Active Filters incorporating a Bridged-T Network, realising Low-Pass and High-Pass Filters for 6th Order Notched Crossovers with f_X = 1kHz

$T_X = T_D$ = 159.2 µs : $(T_N)_{LP} = kT_X$: $(T_N)_{HP} = T_X/k$

All capacitances C1 & C2 are 4.7nF : all resistances in Kohms

k	Filter type	x_N	T_N	x_D	T_D	R1a	R1b	R2	R3	R4
0.6257	LP	0.1600	117.8	0.6723	159.2	40.68	2.109	313.4	33.55	∞
(30dB)	HP	0.1600	215.0	0.6723	159.2	74.23	3.849	571.8	0	693.3
0.6821	LP	0.1400	108.6	0.7313	159.2	29.95	1.709	330.0	34.42	∞
(35dB)	HP	0.1400	233.3	0.7313	159.2	64.37	3.674	709.2	0	617.0
0.7403	LP	0.1000	99.58	0.7801	159.2	21.37	1.115	423.7	33.22	∞
(40dB)	HP	0.1000	254.4	0.7801	159.2	54.59	2.847	1082	0	696.3

The second factor of the sixth order transfer function is produced by active high-pass (with numerators of $s^2T_X^2$) or low-pass filters (with numerators of 1) with denominators $1 + x_D sT_D + s^2T_D^2$, where x_D and T_D are as specified, for example, in Table 4.

The low-pass transfer function

$$F(sT)_{LP} = \frac{1}{1 + x_D sT_D + s^2T_D^2} \qquad - (30)$$

is realised by the circuit of Fig. 3. First component values are chosen for C1 and C2. Then the resistances R1 and R2 are defined as the two values of

$$R1, R2 = [T_D/C2][(x_D/2) \pm \sqrt{\{(x_D/2)^2 - (C2/C1)\}}] \qquad - (31)$$

Note that C2/C1 must be less than $(x_D/2)^2$. The nearer the two ratios are to each other, the more nearly equal will be R1 and R2. Preferably R1 is chosen as the larger.

The high-pass transfer function

$$F(sT)_{HP} = \frac{s^2T_D^2}{1 + x_D sT_D + s^2T_D^2} \qquad - (32)$$

is realised by the circuit of Fig. 4. C1and C2 are chosen preferably as equal values C1. Then

$$R1 = (x_D/2)(T_D/C1) \qquad - (33) \qquad \text{and} \qquad R2 = (2/x_D)(T_D/C1) \qquad - (34)$$

There still remain the transfer functions with the denominators

$$F(sT) = (1 + sT_D)^2 \qquad - (35)$$

These can be realised simply by cascading two CR sections whose CR product is each T_D. In each filter one CR network could be cascaded with the input, the other with the output. Alternatively the second order functions could be realised in the Sallen and Key filters of Figs 3 & 4 with $x_D = 2$, where for both high-pass and low-pass filters C1 is equal to C2 and R1, equal to R2, is T_D / C1.

In this way, each overall sixth-order transfer functions is realised from two or three cascaded active stages

$$F(sT)_{LP} = \frac{1 + qksT_X + k^2s^2T_X^2}{1 + x_6sT_X + s^2T_X^2} * \frac{1}{1 + x_6sT_X + s^2T_X^2} * \frac{1}{1 + 2sT_X + s^2T_X^2} \qquad - (36)$$

and

$$F(sT)_{HP} = \frac{k^2 + qksT_X + s^2T_X^2}{1 + x_6sT_X + s^2T_X^2} * \frac{s^2T_X^2}{1 + x_6sT_X + s^2T_X^2} * \frac{s^2T_X^2}{1 + 2sT_X + s^2T_X^2} \qquad - (37)$$

and the high and low-frequency drivers are connected in opposite polarities. The coefficient q is of course ideally zero.

The addition of signals to produce a seamless, flat, output assumes of course ideal drivers. If the response errors of the higher frequency, tweeter, driver exceed the propensities for forgiveness of the even order crossover, the middle factor of eqn (37) could be substituted by the equalising transfer function

$$F(sT) = \frac{1 + sT_S/Q_T + s^2T_S^2}{1 + x_6sT_X + s^2T_X^2} \qquad - (38)$$

where $T_S = 1/2\Pi f_S$ and f_S is the resonance frequency of the tweeter and Q_T its total Q. This could be realised in an active filter of the same kind as Fig. 2 [2] When this function is cascaded with the transfer function of the driver

$$F(sT) = \frac{s^2T_S^2}{1 + sT_S/Q_T + s^2T_S^2} \qquad - (39)$$

the numerator of eqn (38) cancels with the denominator of eqn (39) to produce the ideal transfer function of the middle factor of eqn. (37).

However, this procedure applies only to crossover functions of sixth or higher order. It must be remembered that the notched crossover, while a sixth order function around the transition frequency, goes to a fourth order slope at extreme frequencies. Thus, because the excursion of a driver rises towards low frequencies at 12dB per octave above its frequency response, its excursion is attenuated only 12dB per octave after such equalisation of a sixth order high-pass notched filter. If a similar procedure was applied to a tweeter with a 4th order crossover function , it would leave little or no protection against excessive excursion at low frequencies. ?

Passive Filters. The fourth order passive filters can be realised using the networks of either Fig. 5 or Fig. 6. Either C3L is parallelled across L2L, as in Fig 5(a) - or L3H across C2H in Fig 5(b) - or L3L is inserted in series with C1L, as in Fig 6(a) - or C3H in series with L1H as in Fig. 6(b). The component values for a low-pass filter of the first kind, in Fig. 5(a), are calculated from the expressions

$$C1L = [3(3 - k^2)/4x_4][T_X/R_O] \qquad - (40)$$

$$C2L = [(1 - 3k^2)/2x_4][T_X/R_O] \qquad - (41)$$

$$C3L = [k^2(3 - k^2)/2x_4(1 - k^2)][T_X/R_O] \qquad - (42)$$

$$L1L = [4x_4/(3 - k^2)]T_X R_O \qquad - (43)$$

$$L2L = [2x_4(1 - k^2)/(3 - k^2)]T_X R_O \qquad - (44)$$

where

$$x_4 = \sqrt{[2(1 - k^2)]} \qquad - (6)$$

The corresponding high-pass components are calculated from the low-pass components, in all cases, using the generalised expressions

$$CnH = T_X^2/LnL \qquad - (45) \qquad \text{and} \qquad LnH = T_X^2/CnL \qquad - (46)$$

In the alternative realisations of the second kind, in Fig 6(a), the low-pass components are

$$C1L = [9(1 - k^2)/4x_4][T_X/R_O] \qquad - (47)$$

$$C2L = T_X/2x_4R_O \qquad - (48)$$

$$L1L = 4x_4 T_X R_O/3 \qquad - (49)$$

$$L2L = 2x_4 T_X R_O/3 \qquad - (50)$$

$$L3L = [4x_4 k^2/9(1 - k^2)]T_X R_O \qquad - (51)$$

while the high-pass component values are again derived from the low-pass values via eqns (45) and (46).

Each version has its uses. In the first kind, Fig. 5(a), C2L goes to zero when $k^2 = 1/3$, i.e. when the following peak height is 30.4dB. Larger values of k cannot be realised, but are unlikely to be needed in practice, with following peak heights higher than -30dB.

The high-pass filter equivalent to this kind of filter, Fig 5 (b), can additionally be adapted to sensitivity control using an auto-transformer [3]. However that network requires high values in the Π network of ~~of~~ inductances transformed from the Π network of capacitances C1L, C2L and C3L, especially L2H, transformed from the small values of C2L. In fact, when k^2 is 1/3, then C2 is zero and L2H goes to infinity. The inductances are more easily realised from a further Δ-Y transformation into the network of Fig. 5(c).

In that network $C1H = [(3 - k^2)/4x_4][T_X/R_O]$ - (52)

$$C2H = [(3 - k^2)/2x_4(1 - k^2)][T_X/R_O] \quad - (53)$$

$$L1H = [4x_4(1 - k^2)(1 - 3k^2)/(3 - k^2)^2]T_X R_O \quad - (54)$$

$$L2H = [6x_4(1 - k^2)/(3 - k^2)]T_X R_O \quad - (55)$$

$$L3H = [4x_4 k^2/(3 - k^2)]T_X R_O \quad - (56)$$

The set of three inductances can be realised either individually or, more conveniently, from two inductors

$$L1H + L2H = [2x_4(1 - k^2)(11 - 9k^2)/(3 - k^2)^2]T_X R_O \quad - (57)$$

$$L1H + L3H = [4x_4(1 - k^2 + 2k^4)/(3 - k^2)^2]T_X R_O \quad - (58)$$

which are wound separately and then coupled together *in series opposing* so that their mutual inductance is L1H, i.e. the coupling coefficient between them is

$$k_{COUPLING} = [2(1 - k^2)(1 - 3k^2)^2 / (1 - k^2 + 2k^4)(11 - 9k^2)]^{1/2} \quad - (59)$$

The resulting filter, Fig. 5(d), may look rather strange but it is eminently practical. The mutual inductance is realised in L1H rather than L3H because that procedure leads to smaller sum inductances L1H + L2H and L1H + L3H over the range of k^2 between 0.333 and 0.157 is are of most practical use.

The coupling coefficients $k_{COUPLING}$ are small enough to be easily achieved. To produce the required coupling, the spacing between the two coils is adjusted until their inductance, measured end to end, is L2H + L31H. The procedure realises all the inductances in the one unit, which can include an auto-transformer[3] and is easily mounted without any worry of stray couplings between individual inductors !

The second version of the low-pass filter, Fig. 6 (a) again needs three inductances, and can again be produced by winding one coil to a value of L1L + L3L another with a value of L2L + L3L and coupling them together *in series opposing* to produce L3L as the mutual inductance between them, as in Fig. 6 (b). This is again produced by varying their coupling until

$$k_{COUPLING} = [2k^4/(3 - 2k^2)(3 - k^2)]^{1/2} \quad - (60)$$

and the inductance end-to-end reads L1L + L2L. Again there is only the one component to mount and no further need to position the inductors to avoid stray coupling. Also in this case, because the mutual inductance L3L is free of a resistive component, the filter is capable of a better null.

The high pass filter of the second kind, Fig 6 (c) is less desirable than the first kind. It requires three capacitors, one of which C3 is comparatively large. Component values for a crossover frequency of 1000Hz and a terminating resistance of 10 ohms are presented in Table 5 for all four realisations of each of the three fourth order versions, with following peaks of approximately -30dB, -35dB and -40dB.

Table 5. Fourth Order Passive Notched Crossovers.
Component Values for $f_X = 1000$Hz and $R_0 = 10$ohms

Low-Pass Filter (with C3 in parallel with L2)

k	*L1(μH)*	*C1(μF)*	*L2(μF)*	*C3(μF)*	*C2(μF)*
0.5774	2757	27.57	919	9.189	0
0.5000	2835	26.80	1063	5.956	1.624
0.4472	2876	26.42	1150	4.404	2.516
0	3001	25.32	1501	0	5.627

High-Pass Filter (with L1 L2 & L3 in network around C2)

k	*C1(μF)*	*L1(μH)*	*C2(μF)*	*L3(μH)*	*L2(μH)*	$k_{COUPLING}$
0.5774	9.189	0	27.57	918.9	2757	0
0.5000	8.934	193.3	23.82	708.8	3190	0.1107
0.4472	8.808	328.7	22.02	575.2	3451	0.1778
0	8.440	1000.3	16.88	0	4502	0.4264

Low-Pass Filter (with L3 in series with C1)

k	*L1(μH)*	*C1(μF)*	*L3(μH)*	*L2(μH)*	*C2(μF)*	$k_{COUPLING}$
0.5774	2450	20.68	408.4	1225	6.892	0.1890
0.5000	2599	21.93	288.8	1299	6.497	0.1348
0.4472	2684	22.65	223.7	1342	6.291	0.1048
0	3001	25.32	0	1501	5.627	0

High-Pass Filter(with C3 in series with L1)

k	*C1(μF)*	*L1(μH)*	*C3(μF)*	*C2(μF)*	*L2(μF)*
0.5774	10.34	1225	62.02	20.68	3676
0.5000	9.746	1155	87.72	19.49	3898
0.4472	9.437	1118	113.2	18.87	4026
0	8.440	1000	∞	16.88	4502

Input Impedance. The input impedances of the passive filters are identical for the two kinds of realisation in Figs 5 and 6. They are generally, and not surprisingly, similar to those of the Linkwitz-Riley filters [4]. The high-pass and low-pass filters present individually resistive and reactive components that are symmetrical in frequency in that their magnitudes for the high-pass filter at any frequency nf_X are the same as for the low-pass filter at the frequency f_X/n . The sign of the reactive components is always negative for the high-pass filter and always positive for the low-pass filter but their magnitudes are equal, and cancel in parallel, only at the transition frequency. At other frequencies, the magnitude of their combined reactance is never less than 3 times the nominal, terminating, impedance R_0 .

The resistive component of each filter is 4Ro at the transition frequency, (the two in parallel present 2Ro), rising rapidly at frequencies outside the pass-band. Within the pass-band the resistive component diminishes through Ro at the notch frequency of the other filter to a minimum (never lower than 0.96Ro) before returning to Ro at extreme frequencies. The reason is that, as explained earlier, each filter must, at frequencies in its pass-band beyond the notch of the other filter, produce an output a little greater (0.27dB maximum) than its input so as to maintain a flat combined output. To produce more power from a low(virtually zero) impedance source, the filter must present a lower resistive component of input impedance.

Like most passive crossovers, these networks require ideally an accurate and purely resistive termination. Unless the driver presents a good approximation to such a resistance, its input will need to be shunted by an appropriate impedance correcting network[5].

Conclusion. The notched crossover systems, especially those using even order functions, offer improvements in performance, particularly when rapid attenuation is needed close to the transition frequency. Although their performance in lobing with non-coincident drivers has not been examined specifically, it is expected to be similar to that of the Linkwitz-Riley crossovers, because their two outputs maintain a constant zero phase difference across the transition.

The passive filters that utilise coupling between inductors also offer convenience in realisation and in mounting in the cabinet as a single unit.

The odd-order functions, whose high-pass and low-pass outputs add in quadrature, have been included for completeness, though they would seem to be of less general interest than those of even order.

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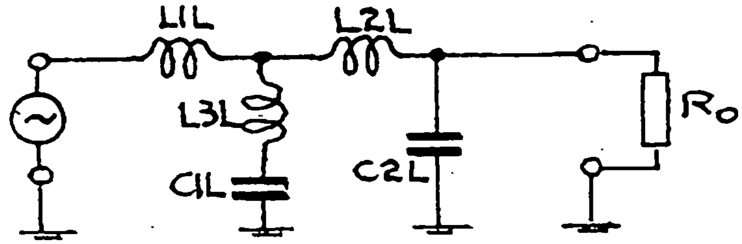

Fig 6(a) Passive fourth-order low-pass filter (second kind)

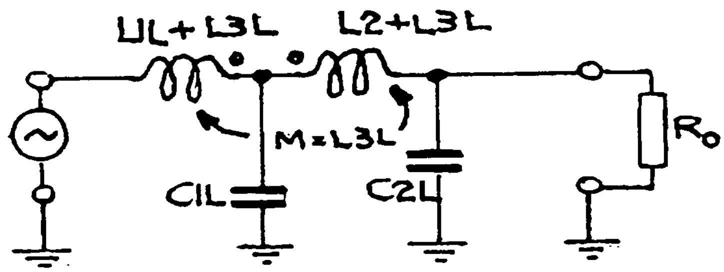

Fig 6(b) Passive fourth-order low-pass filter (second kind)
with inductances of Fig 6(a) realised as a coupled pair (series opposing)

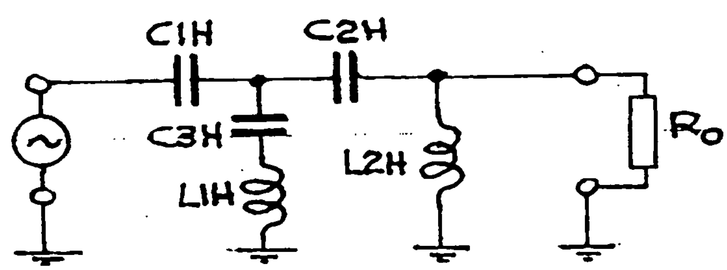

Fig 6(c) Passive fourth-order high-pass filter (second kind)

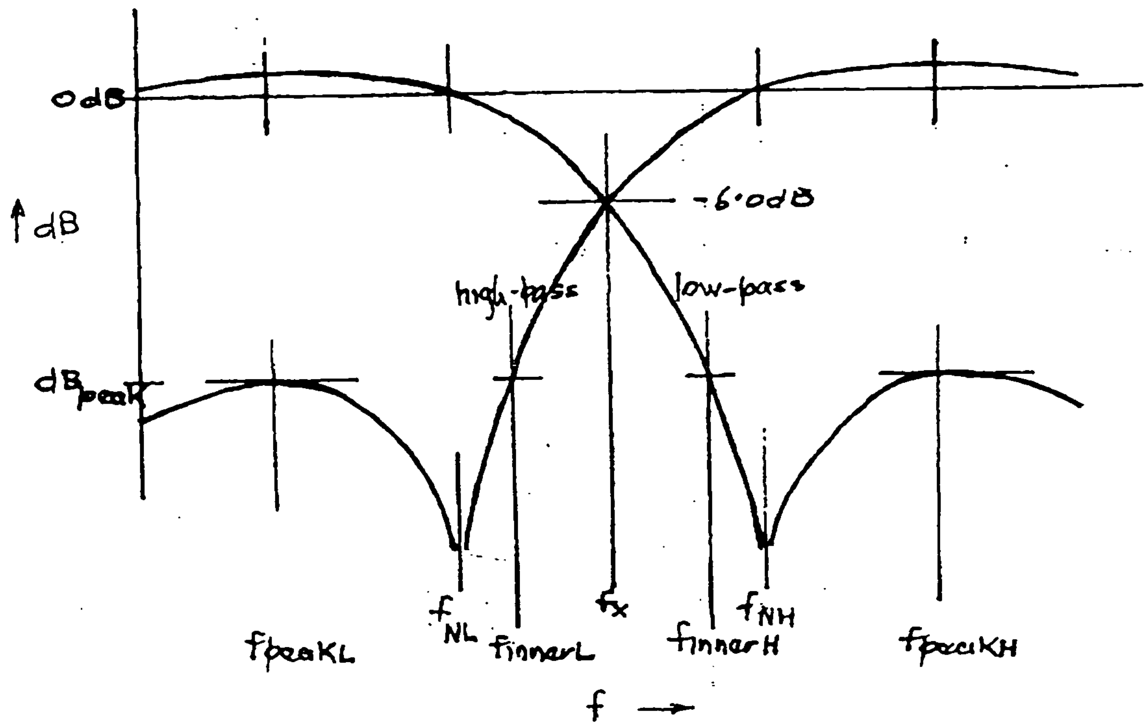

Fig 1. Generalised responses of even order notched crossovers

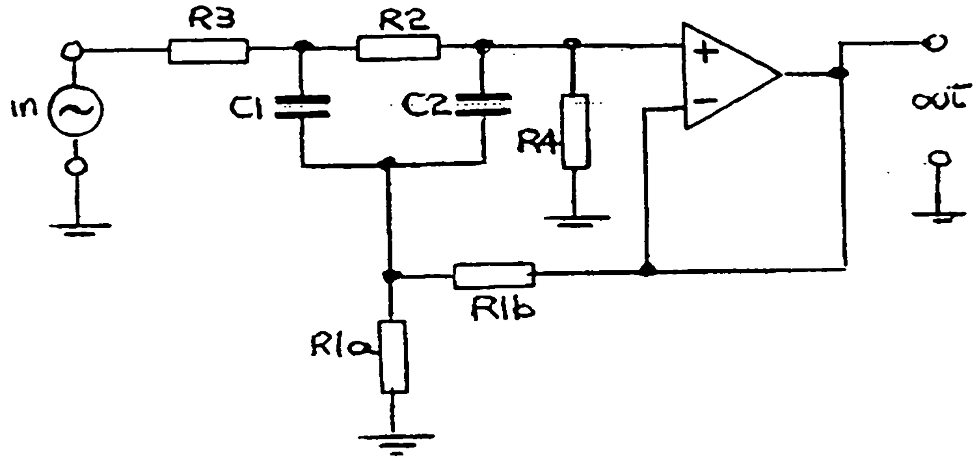

Fig 2. Sallen & Key active filter incorporating a bridged-T network

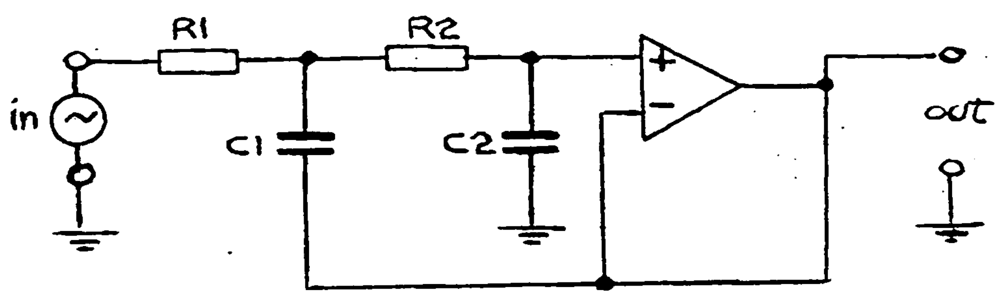

Fig 3 Sallen & Key active low-pass filter

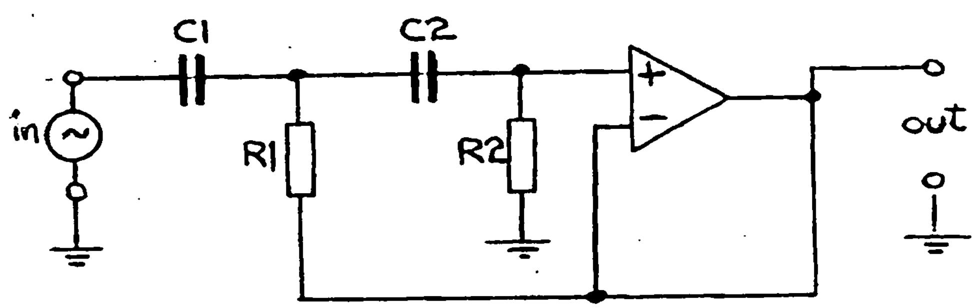

Fig 4 Sallen & Key active high-pass filter

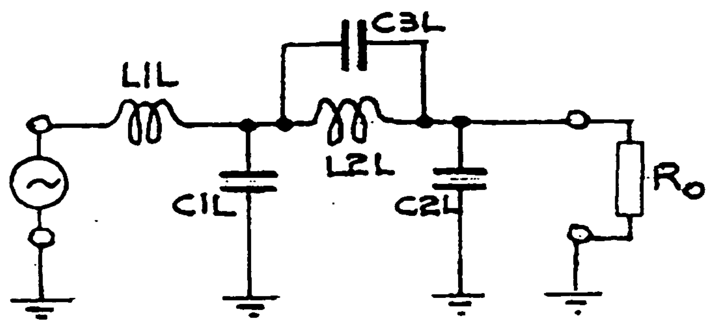

Fig 5(a) Passive fourth-order low-pass filter (first kind)

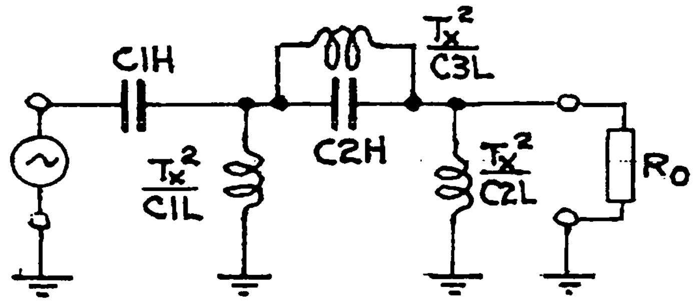

Fig 5(b) Passive fourth-order high-pass filter (first kind) with components transformed, $C_{nH} = T_X^2 / L_{nL}$ & $L_{nH} = T_X^2 / C_{nL}$, from Fig 5(a)

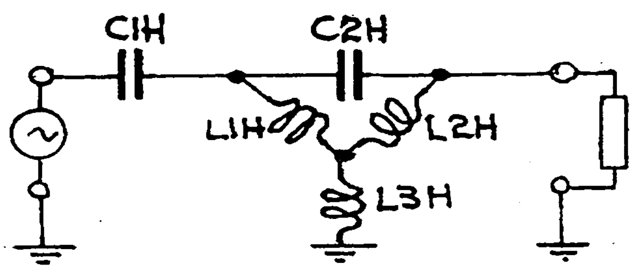

Fig 5(c) Passive fourth-order high-pass filter (first kind)
with inductances the result of Δ-Y transformation from Fig 5(b)

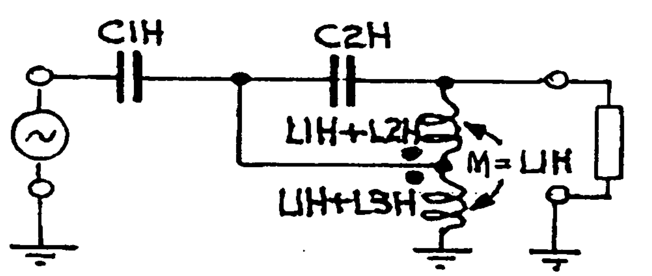

Fig 5(d) Passive fourth-order high-pass filter (first kind)
with inductances of Fig 5(c) realised as a coupled pair (series opposing)

List of Captions.

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Fig 4 Sallen & Key active high-pass filter

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Fig 5(b) Passive fourth-order high-pass filter (first kind)
with components transformed $CnH = T_X^2 / LnL$ & $LnH = T_X^2 / CnL$ from Fig 5(a)

Fig 5(c) Passive fourth-order high-pass filter (first kind)
with inductances the result of Δ-Y transformation from Fig 5(b)

Fig 5(d) Passive fourth-order high-pass filter (first kind)
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